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Demodulation Type Single-Phase PLL with DC Offset Rejection

H. Ahmed and M. Benbouzid

This letter proposes demodulation type PLL for phase and frequency estimation of single-phase system that can reject DC offset. Using results from the adaptive estimation literature, this letter proposes a linear parametric model-based initial phase angle estimation approach. Then by using differentiation and integration operation on the estimated initial phase angle, the frequency is estimated. This avoids the use of any low-pass filter unlike conventional demodulation-based technique. Moreover, unlike existing demodulation-based technique, the proposed technique can completely reject DC offset. Comparative experimental results, provided with state of the art DC offset rejection-enhanced phase locked-loop (EPLL), clearly demonstrate the suitability of the proposed technique.

1. Introduction: Phase and frequency play a very important role in various control, monitoring, protection, and quality assessment of power system and grid-connected converters (GCC). Some potential applications are: grid synchronization of GCC [1, 2, 3], control of PWM rectifier [4], grid monitoring [5, 6] etc. to name a few. All these applications require fast and accurate estimation of phase and frequency.

In this letter our focus is on phase and frequency estimation for single-phase system. Some of the existing techniques for single-phase systems are: Kalman filter [7], adaptive observer [8], phase-locked loop (PLL) [9, 10, 11, 12], enhanced PLL (EPLL) [13, 14], adaptive filter-based frequency-locked loop (FLL) [15], and demodulation-based technique [16, 17].

Kalman filter works very well in noisy environment. It can provide fast and accurate estimation. However, the performance and stability depend on tuning the process and measurement noise parameters. In addition, initial conditions also play a great role. Kalman filter requires PLL/FLL to estimate the frequency. As such the tuning of the whole system can be complicated. Adaptive observer [8] uses the same dynamical model as Kalman filter, however, from a deterministic framework. Adaptive observer is sensitive to unmodeled dynamics. Moreover, performance degrades in noisy environment.

PLL [9, 10, 11, 12] is undoubtedly the most popular technique available in the literature. Many existing PLL results for single-phase system use the idea of synchronous reference frame - PLL (SRF-PLL) originally proposed for three-phase system. To apply SRF-PLL in single-phase system, a virtual orthogonal signal needs to be generated to implement Clarke to Park transformation. Tuning the parameters of PLL requires a trade-off between fast dynamic response and estimation accuracy. Enhanced PLL [13, 14] can be considered as a nonlinear version of the PLL. FLLs [15] also require orthogonal signal to be generated. Then using the opposite phase relationship between the input variables, the frequency can be easily estimated.

Traditional demodulation-based frequency estimation technique [16] is very simple. By multiplying fixed frequency signals with the grid voltage signal, the demodulated voltages can be obtained. Then by using high order low-pass filter (LPF) tuned at two times the fundamental frequency, the frequency can be estimated. LPF's are non-ideal in nature. In nominal operating condition, this may not be an issue. However, during off-normal operating conditions, LPFs may introduce large attenuation of the filtered signal's amplitude due to high cut-off frequency. To overcome this issue, frequency adaptive quadrature oscillator-based demodulation technique is reported in [17]. An advantage of [17] is that the cut-off frequencies are significantly lower than standard demodulation-based technique. Moreover, the presence of two LPFs also provide additional robustness. However, these benefits come at the cost of higher computational complexity. One important limitation of the conventional and modified demodulation-based techniques are that they can't handle DC offset explicitly. DC offset may arise from signal conversion by analog-to-digital converter, transformer saturation etc.

In this letter, we focus on demodulation-based technique. The objective is to propose an improved estimator that can reject DC offset and similarly works to demodulation based-technique. The proposed estimator uses results from the adaptive estimation literature [18] to estimate the initial phase angle of the grid voltage signal. Then by using differentiation and

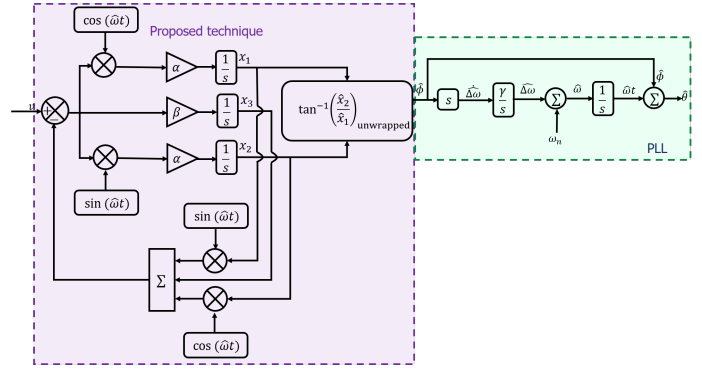


Fig. 1. Block diagram of the proposed DT-PLL.

integration operation on the estimated phase angle, the frequency and instantaneous phase are estimated. Comparative experimental results are provided with DC offset rejection-based EPLL [13].

The rest of this letter is organized as follows: Sec. 2 describes the conventional demodulation-based technique and the proposed technique while Sec. 3 deals with the comparative experimental results. Finally, Sec. 4 concludes this letter.

2. Unknown Parameter Estimation:

(a) Conventional Demodulation-based Technique

A single-phase grid voltage signal is given by:

$$v(t) = A \sin(\underbrace{\omega t + \phi}_{\theta}) + A_0 \quad (1)$$

where A is the amplitude, $\omega = \omega_n + \Delta\omega$ is the angular frequency (with $\omega_n = 100\pi$ being the nominal angular frequency and $\Delta\omega$ is the frequency deviation), ϕ is the phase, $\theta \in [0, 2\pi)$ is the instantaneous phase, and A_0 is the DC offset. To estimate the unknown parameters ω and θ , conventional demodulation-based techniques [16] work by estimating the phase, $\varphi = \Delta\omega t + \phi$. Assuming $A_0 = 0$, conventional demodulation technique is summarized below:

From the grid signal $v(t)$, the demodulated voltages can be obtained as follow:

$$v_c(t) = v(t) \cos(\omega_n t) = \frac{A}{2} \sin(\varphi) + \frac{A}{2} \sin(2\omega_n t + \varphi) \quad (2)$$

$$v_s(t) = v(t) \sin(\omega_n t) = \frac{A}{2} \cos(\varphi) - \frac{A}{2} \cos(2\omega_n t + \varphi) \quad (3)$$

As can be seen from Eq. (2) and (3), demodulated voltages contain double fundamental frequency component which can be easily removed using a suitably designed LPF. From the filtered demodulated voltages, the phase can be estimated as:

$$\hat{\varphi} = \arctan(v_{cf}/v_{sf})_{\text{unwrapped}} \quad (4)$$

where the subscript f indicates filtered signal and unwrapped indicates that the estimated phase is wrapped. Then the angular frequency can be estimated as :

$$\hat{\omega} = \omega_n + \dot{\hat{\varphi}} \quad (5)$$

Conventional demodulation-based technique requires two high order LPFs with the cut-off frequency being two times the fundamental frequency. LPFs have non-ideal characteristics. As such, large attenuation of the amplitude during large frequency deviation may not be avoided. To overcome this limitation, Reza *et al.* [17] proposed quadrature oscillator-based frequency adaptive demodulation technique. This technique still requires two low-pass filter, however, the cut-off frequency is lower than the fundamental frequency. This overcomes the amplitude attenuation problem, however, comes at the cost of computational complexity. Moreover, neither the conventional nor the modified technique can handle DC offset.

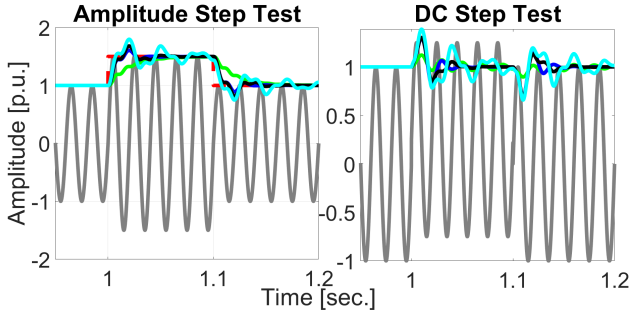


Fig. 2 Simulation results for parameter tuning. Grid signal (—); amplitude (—); $\alpha = 100, \beta = 12.5$ (—); $\alpha = 400, \beta = 50$ (—); $\alpha = 500, \beta = 100$ (—); $\alpha = 1000, \beta = 200$ (—).

(b) *Proposed Technique*

Using trigonometric manipulation, the grid voltage signal (1) can be written as:

$$v(t) = \sin(\omega t)A \cos(\phi) + \cos(\omega t)A \sin(\phi) + A_0$$

$$v(t) = \underbrace{\begin{bmatrix} \sin(\omega t) & \cos(\omega t) & 1 \end{bmatrix}}_{C^T} \underbrace{\begin{bmatrix} A \cos(\phi) \\ A \sin(\phi) \\ A_0 \end{bmatrix}}_{\mathbf{x}} \quad (6)$$

where $\mathbf{x} = [x_1 \ x_2 \ x_3]^T$ is the parameter vector. Eq. (6) is a linear parametric model provided that ω is constant. In practice, ω is not constant, as such, ω needs to be estimated. This will be discussed later on. Assuming that $\hat{\omega}$ is available, then the estimator adapted from [18, Ch. 4] can be applied to estimate the parameters:

$$\dot{\hat{x}}_1 = \alpha \sin(\hat{\omega} t)(v - \hat{v}) \quad (7a)$$

$$\dot{\hat{x}}_2 = \alpha \cos(\hat{\omega} t)(v - \hat{v}) \quad (7b)$$

$$\dot{\hat{x}}_3 = \beta(v - \hat{v}) \quad (7c)$$

$$\hat{v} = \hat{C}^T \hat{\mathbf{x}} \quad (7d)$$

where $\alpha, \beta > 0$ are tuning gains. In Eq. (7), time argument is avoided for notational simplicity. As shown in [18], an advantage of estimator (7) is that it provides exponential convergence of the estimation errors. Similar to demodulation-based technique, the estimator (7) also use multiplication of trigonometric function to extract the phase. However, unlike demodulation-based techniques, no low-pass filtering is required. Moreover, the estimator (7) can be used to generate quadrature signal from v as well. Quadrature signal of v i.e. $v_Q = A \cos(\omega t + \phi)$ can be estimated as

$$\hat{v}_Q = \cos(\hat{\omega} t)\hat{x}_1 + \sin(\hat{\omega} t)\hat{x}_2 \quad (8)$$

Estimator (7) has the following transfer function:

$$\hat{v}(s) = v(s)/(\tau s + 1) \quad (9)$$

where $\tau = 1/(\alpha + \beta)$. Transfer function (9) shows that low-pass filtering is an inherent property of the estimator. This can be very useful in practical application.

From the estimated parameter vector $\hat{\mathbf{x}}$, the phase of the grid voltage signal can be extracted as:

$$\hat{\phi} = \arctan(\hat{x}_2/\hat{x}_1)_{\text{unwrapped}} \quad (10)$$

In the nominal condition, $\hat{\phi}$ represents only the phase angle. However, during frequency variation, in addition to the phase angle, $\hat{\phi}$ also contains the deviation from the nominal frequency. As a result, the time derivative of $\hat{\phi}$ can be approximated as the frequency deviation dynamics i.e. $\dot{\hat{\omega}} = \dot{\hat{\phi}}$. Then passing this signal through an integrator with gain γ , the estimated frequency deviation $\Delta\omega$ can be easily obtained. Block diagram of the proposed demodulation type PLL (DT-PLL) is given in Fig. 1.

Due to the presence of trigonometric quantities in the estimator dynamics (7), small-signal modeling-based parameter tuning can be

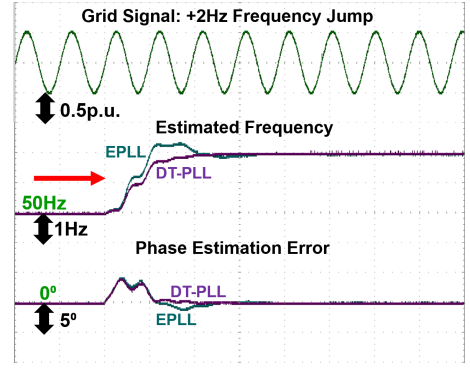


Fig. 3. Experimental results for Test-I.

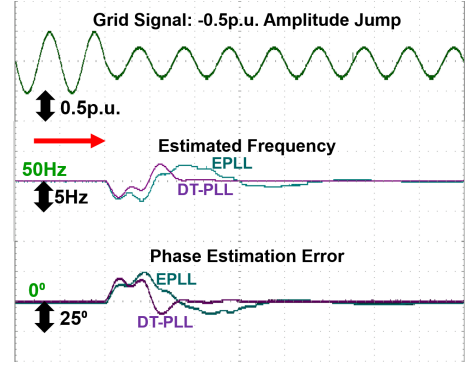


Fig. 4. Experimental results for Test-II.

complicated for DT-PLL. Simulation studies can be very useful in this regard. To tune the estimator parameters, first we set the frequency estimation gain as $\gamma = 50$. Then two simulation tests have been considered. In the first case, amplitude jump of $+0.5p.u.$ is considered while in the second test DC offset jump of $+0.15p.u.$ is considered. Results of the simulation are given in Fig. 2. This figure shows the estimated amplitudes for four different values of α and β . For smaller values of α and β , the convergence speed can be observed to be very slow while high value of α and β leads to oscillation in the estimated amplitude. $\alpha = 400$ and $\beta = 50$ found to be a compromise between convergence speed and peak overshoot. As such this values are selected. Frequency estimation speed is controlled by γ . Through extensive numerical simulation, $\gamma = 50$ was found to be a good value.

3. Results and Discussions: To demonstrate the effectiveness of the proposed DT-PLL, dSPACE 1104 board-based experimental study is considered. As a comparison tool, we have selected DC offset rejection-based enhanced PLL (EPLL) [13]. Both techniques are implemented in Matlab/Simulink for code generation purpose with sampling frequency 10KHz. EPLL parameters are selected as: $\mu_0 = 85$, $\mu_1 = \mu_3 = \omega_n$, and $\mu_2 = 30000$. Parameters values are taken from [13]. DT-PLL parameters are selected as: $\alpha = 400$, $\beta = 50$, and $k = 50$. In all the figures presented in this Section, X-axis indicates 20msec./Div.

Test I +2Hz Frequency jump: Fig. 3 shows the actual grid voltage, estimated frequency and the phase estimation error. From Fig. 3, it can be observed that the the estimated frequency by DT-PLL converged very fast in ≈ 2.5 cycles without any overshoot while EPLL took ≈ 3 cycles with around 0.3Hz overshoot. Similarly the phase estimation error converged in ≈ 2 cycles for DT-PLL while EPLL took ≈ 2.5 cycles. These results show that DT-PLL is very effective in frequency tracking. It is to be mentioned here that many PLL can provide ≈ 2 cycle convergence. However, those results are in general obtained without any DC bias estimation loop. DC bias estimation loop slows down the dynamic response. As such the convergence time of DT-PLL is competitive w.r.t. existing literature.

Test II -0.5p.u. Amplitude jump: In this test, large amplitude jump is considered. In the context of low voltage ride through (LVRT) capability, this test is very significant. As shown in Fig. 4, both techniques reacted very fast to the amplitude jump. However, DT-PLL converged significantly faster than EPLL. Both techniques don't employ any explicit

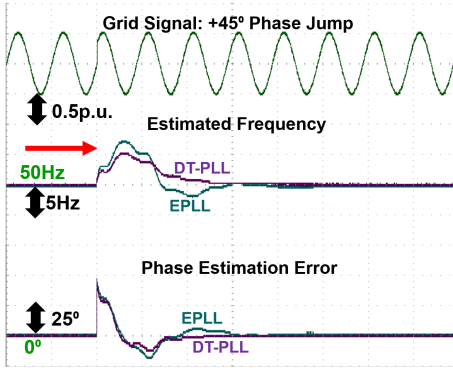


Fig. 5. Experimental results for Test-III.

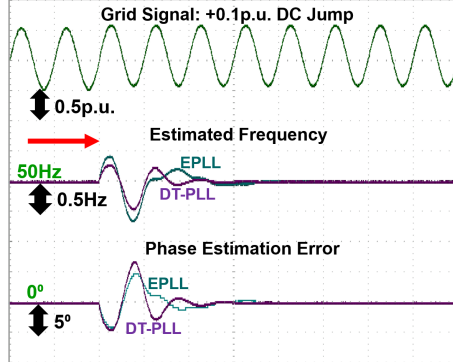


Fig. 6. Experimental results for Test-IV.

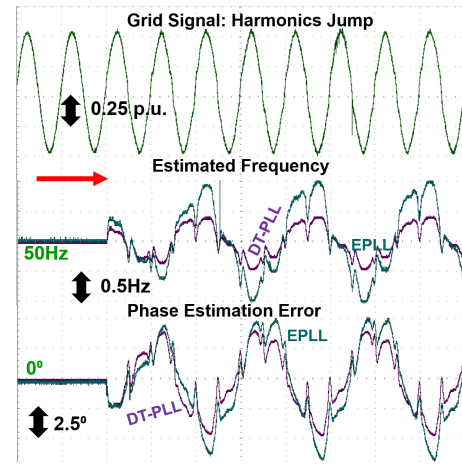


Fig. 7. Experimental results for the harmonics robustness test.

gain normalization technique. However, experimental results show that DT-PLL is significantly less sensitive to voltage sag.

Test III +45° Phase jump: Fig. 5 shows comparative results for phase jump test. This figure demonstrates that DT-PLL has significantly less peak overshoot compared to EPLL. Moreover, fast convergence can also be observed for DT-PLL.

Test IV +0.1 p.u. DC jump: Fig. 6 shows the results for DC offset jump test. Experimental results show that both techniques have similar peak overshoot. However, DT-PLL converged $\approx 0.5 \sim 1$ cycle faster than EPLL.

Test V Harmonics robustness: In this test, suddenly the grid voltage is corrupted with harmonics. Considered harmonics components are: 3rd – 1.3%, 5th – 2.9%, 7th – 1.7%, 9th – 2.6%, 11th – 1.8%, 13th – 1%, sub-harmonics 30Hz – 3.1%, and inter-harmonics 170Hz – 2.7%. Fig. 7 shows the results for harmonics robustness test. Experimental results confirm that the proposed PLL has higher harmonic filtering capability than that of EPLL. Proposed technique has a ripple magnitude of $\approx \pm 0.4$ Hz and $\approx \pm 3.2^\circ$ while EPLL has a ripple magnitude of $\approx \pm 1$ Hz and $\approx \pm 5^\circ$. As such the proposed technique can be claimed to be very suitable even in distorted grid condition.

4. Conclusion: A demodulation type single-phase PLL has been proposed in this letter to estimate the phase and frequency of single-phase system. First, a linear parametric model has been developed. Then a simple estimator is proposed to estimate the initial phase angle similar to conventional demodulation-based technique. Using the initial phase angle, a simple frequency tracker is proposed. Comparative experimental results with enhanced phase locked-loop demonstrated the effectiveness of the proposed demodulation type PLL.

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